DESIGN AND DEVELOPMENT OF MICROWAVE AMPLIFIER

by

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INDIAN INSTITUTE OF TECHNOLOGY KANPUR

July, 1997

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A Thesis Submitted

in Partial Fulfilment of the Requirements

for the Degree of

PGDIIT

by

SALADI VALIBABU

to the

DEPARTMENT OF ELECTRICAL ENGINEERING

INDIAN INSTITUTE OF TECHNOLOGY KANPUR

July, 1997

EE-1997-M-VAL-DES

CERTIFICATE

It is certified that the work contained in the thesis entitled "Design and Development of Microwave Amplifier", by Saladi Valibabu has been carried out under my supervision and that this work has not been submitted elsewhere for a degree.

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23 July, 1997

ABSTRACT

In this thesis, Design and development of MICROWAVE AMPLIFIER has been presented. Design techniques in various cases, like unilateral and bilateral in unconditionally stable and potentially unstable cases, have been studied.

Software has been developed for complete design of a Narrowband microwave transistor amplifier. The software selects a point in source reflection plane in such a way that it should satisfy specified gain and noise figure. For the given parameters, if there is no point in Γ_S-plane, which satisfies both gain and noise figure, then the software will choose Γ_S, which satisfies either specified gain or specified noise figure and nearest possible noise figure or gain to the given values respectively. For the so chosen Γ_S, load reflection coefficient is calculated. Matching networks are designed to match the device to source and load. In the matching networks design, software uses single stub matching technique for narrowband purpose. Software also makes calculation of microstrip line parameters for the designed matching networks and gives datafile, from which layout can be prepared.

By using the software a narrowband microwave transistor amplifier has been designed. Layout was made on copper coated Duroid ($\epsilon_r = 2.23$) sheet. Components have been soldered and layout was fixed in a metal enclosure.

Alignment of the amplifier is carried out by balanced shunt stub method. Various measurement tests have been done and the values of gain and noise figure were noted down at various frequencies. The Amplifier works satisfactorily with a gain of 14.1 dB and noise figure of 2.3 dB for bandwidth of approximately 0.103 GHz.

ACKNOLEDGEMENTS

I express my deep sense of gratitude to my thesis supervisor Dr. Animesh Biswas for his invaluable active guidance , inspiration and constant encouragement throughout my thesis work. His guidance at every stage of my cource of thesis work not only led to enormous improvement in the content of this thesis but also gave a practice insight into the project work for which I am very grateful to him.

I am very thankful to Dr. Apu Sivadas, Dr. Harish, Mr. Pavan Kumar, Mr. R. K. Joshi for their help, innumerable comments and suggestions.

I wish to thank my batchmates Rajibabu, Time, Sugadhan, Pradeep, Gupta, Archana and Vardhini always stood by me during my cource work.

I am, forever, thankful to my parents for their blessings and encouragement received at every juncture, which has helped me to complete my work smoothly.

I wish to express my gratitude to the Dept. of DOORDARSHAN, Govt. of India, for providing me an opportunity to pursue this P. G. cource.

SALADI VALIBABU

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Chapter 1

Introduction

1.1 Introduction

Microwave systems are extensively used in the milatary like radar, navigation, surveillance, weapon guidance systems and commercially like cellular phones, data communication links, microwave ovens and video/audio communications.

Microwave systems consists of basic circuits like filters, amplifiers, oscillators and mixers. In the design analysis of a microwave amplifier knowledge of the S-parameters , matching networks, transmission line theory and signal flow graph analysis are required. These four topics are discussed in this chapter.

1.2 Microstrip matching networks

Microstrip lines are used extensively in building microwave transistor amplifiers because of their ease in fabrication using printed circuit techniques. Network inter-connections

and placement of lumped active and passive devices are easily made on its metal surface of the printed circuit boards(PCB).

A microstrip line is by definition a transmission line consisting of a strip conductor and a ground plane separated by a dielectric medium. Figure 1.1 illustrates the microstrip geometry. The dielectric material serves as a substrate and it is sandwitched between the strip conductor and the ground plane. Some typical dielectric substrates are Duroid ($\epsilon_r = 2.23$), quartz($\epsilon_r = 3.78$),

alumina($\epsilon_r = 9.7$), and silicon($\epsilon_r = 11.7$)

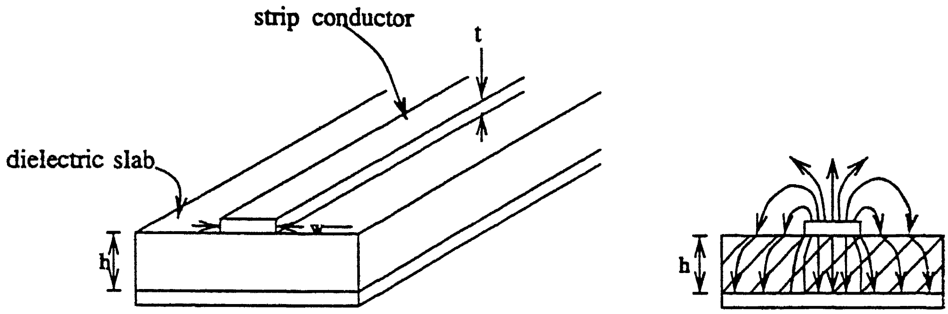

Figure 1.1: Microstrip geometry and field configurations

The electromagnetic field lines in the microstrip are not contained entirely in the substrate. Therefore,the propagating mode in the microstrip is not a pure transverse electromagnetic mode(TEM mode) but a quasi-TEM. Assuming a quasi -TEM mode of propagation in the microstrip line, the phase velocity is given by expression 1.1 [1]

$$v_p = \frac{c}{\sqrt{\epsilon_{ff}}} \tag{1.1}$$

where c is speed of light (i,e $3 \times 10^8 m/s$) and ϵ_{ff} is relative permittivity of the dielectric substrate.

The characteristic impedance of microstrip line is given by expression 1.2

$$Z_0 = \sqrt{\frac{L}{C}} = v_p L = \frac{1}{v_p C} \tag{1.2}$$

where L and C are the inductance and capacitance per unit length respectively and v_p is phase velocity given by

$$v_p = \frac{1}{\sqrt{LC}} \tag{1.3}$$

The effective microstrip permittivity is defined as the ratio of the capacitance of microstrip line with dielectric to the capacitance of the same microstrip line without dielectric(i,e with air).

$$\epsilon_{eff} = \frac{C}{C_a} = \left(\frac{C}{v_p}\right)^2 \tag{1.4}$$

In microstrip circuit design, the calculated length is normally expressed in terms of λ_g. Once with λ_g obtained $\left(\lambda_g = \frac{\lambda_0}{\sqrt{\epsilon_{eff}}}\right)$, the physical length can be easily found.

Formula for synthesis and analysis of microstrips:

There are many methods to calculate the two static capacitances with and with out air. In this section, the results of one such method are given.

Synthesis formulae

Synthesis means finding w/h and ϵ_{eff} with the knowledge of the values of Z_0 and ϵ_r [2,3]

For narrow strips ($Z_0 > (44 - 2\epsilon_r)$ohms)

$$\frac{w}{h} = \left(\frac{e^H}{8} - \frac{1}{4e^H}\right)^{-1} \tag{1.5}$$

where,

$$H = \frac{Z_0\sqrt{2(\epsilon_r + 1)}}{119.9} + \frac{1}{2}\frac{\epsilon_r - 1}{\epsilon_r + 1}\left(\ln\frac{\pi}{2} + \frac{1}{\epsilon_r}\ln\frac{4}{\pi}\right) \tag{1.6}$$

and,

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2}\left[1 - \frac{1}{2H}\frac{\epsilon_r - 1}{\epsilon_r + 1}\left(\ln\frac{\pi}{2} + \frac{1}{\epsilon_r}\ln\frac{4}{\pi}\right)\right]^{-2} \tag{1.7}$$

For wide strips ($Z_0 > (44 - 2\epsilon_r)$ohms)

$$\frac{w}{h} = \frac{2}{\pi}\left[(d_\epsilon - 1) - \ln(2d_\epsilon - 1)\right] + \frac{\epsilon_r - 1}{\pi\epsilon_r}\left[\ln(d_\epsilon - 1) + 0.293 - \frac{0.517}{\epsilon_r}\right] \tag{1.8}$$

where,

$$d_\epsilon = \frac{59.95\pi^2}{Z_0\sqrt{\epsilon_r}} \tag{1.9}$$

and,

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\left(1 + 10\frac{h}{w}\right)^{-0.555} \tag{1.10}$$

Analysis formulae

for wide strips,

$$Z_0 = \frac{119.9\pi}{2\sqrt{\epsilon_r}}\left\{\frac{w}{2h} + \frac{\ln 4}{\pi} + \frac{\ln(e\pi^2/16)}{2\pi}\left(\frac{\epsilon_r - 1}{\epsilon_r^2}\right) + \frac{\epsilon_r + 1}{2\pi\epsilon_r}\left[\ln\frac{\pi e}{2} + \ln\left(\frac{w}{2h} + 0.94\right)\right]\right\}^{-1} \tag{1.11}$$

for narrow strips,

$$Z_0 = \frac{119.9}{\sqrt{2(\epsilon_r + 1)}} \left\{ \ln \left[\frac{4h}{w} + \sqrt{16\left(\frac{h}{w}\right)^2 + 2} \right] - \frac{1}{2}\frac{\epsilon_r - 1}{\epsilon_r + 1} \left(\ln \frac{\pi}{2} + \frac{1}{\epsilon_r} \ln \frac{4}{\pi} \right) \right\} \tag{1.14}$$

1.3 Matching networks

As the operating frequency increases beyond a few hundred mega hertzs, the value of capacitance and inductance so calculated may become impractically small. Then distributed elements have to be employed instead. The realization of distributed elements is based on the impedance transforming properties of transmission line. The most readily available structure in the high frequency domain, especially for the circuits involving active devices, is open microstrip.

There are two methods, which are commonly used for matching transmission lines. They are:

1. Stub matching
   - (a) Single stub matching
   - (b) Double stub matching
2. Quarter wave transformer

Generally, Smith charts are used to design these matching networks. An open circuited or short circuited stub can easily be realised in microstrip configuration and can transform a 50Ω resistance into any value of impedance. Also, a quarter wave microstrip line can be used to change a 50Ω resistor to any value of resistance.

To minimise transition interaction between the shunt stubs and series transmission lines, two parallel shunt stubs should be used in balanced configuration along the main line and must provide the same admittance as single stub. Therefore, the admittance of each side of the balanced stub must be equal to half of the total admittance.

1.4 Scattering Parameters

In the theory of network analysis,it is well known that a network, can be completely specified by a set of four parameters i,e y-,z-, h- or ABCD parameters [4]. All these network characterizations are based on the total voltages and currents appearing on the terminals of the two ports.

In high frequency applications the above said parameters are seldom used because of the following problems:

1. Total voltages and currents are difficult to measure.
2. In the measurements of these two-port parameters, short and open circuits are required. However, they are difficult to realise over a broadband of frequencies
3. Furthermore, most active devices or circuits are not open- or short- circuit stable.

A two-port network can be completely described by two equations 1.13 and 1.14 via two-port parameters set such as the y-parameter i,e [4]

$$I_1 = y_{11}V_1 + y_{12}V_2 \tag{1.13}$$

$$I_2 = y_{21}V_1 + y_{22}V_2 \tag{1.14}$$

where V_1 and I_1 are total input terminal quantities and V_2 and I_2 are total output terminal quantities.

The quantities V_1, I_1, V_2 and I_2 can be expressed in terms of travelling voltage and current waves as

$$V_1 = V_1^+ + V_1^- \tag{1.15}$$

$$I_1 = \frac{V_1^+ - V_1^-}{Z_0} \tag{1.16}$$

$$V_2 = V_2^+ + V_2^- \tag{1.17}$$

$$I_2 = \frac{V_2^+ - V_2^-}{Z_0} \tag{1.18}$$

where + and - superscripts refer to whether the travelling wave is going into or coming out from the two port network and Z_0 is the characteristic impedance of the system.

On substituting expressions 1.13 and 1.14 in the expressions 1.15 to 1.18, we obtain

$$V_1^- = f_{11}(y, Z_0)V_1^+ + f_{12}(y, Z_0)V_2^+ \tag{1.19}$$

$$V_2^- = f_{21}(y, Z_0)V_1^+ + f_{22}(y, Z_0)V_2^+ \tag{1.20}$$

In equations 1.19and 1.20, f_{ij} s are functions of y-parameters (or z or ABCD or h) and the impedance level Z_0 chosen.

Equation 1.19 and 1.20 will not change if we divide throughout by $\sqrt{Z_0}$ and define

$$a_1 = \frac{V_1^+}{\sqrt{Z_0}} \tag{1.21}$$

... square root of the power incident at port 1.

$$a_2 = \frac{V_2^+}{\sqrt{Z_0}} \tag{1.22}$$

... square root of the power incident at port 2.

$$b_1 = \frac{V_1^-}{\sqrt{Z_0}} \tag{1.23}$$

... square root of the power emitted at port 1.

$$b_2 = \frac{V_2^-}{\sqrt{Z_0}} \tag{1.24}$$

... square root of the power emitted at port 2. and $S_{11} = f_{11}(y, Z_0), S_{12} = f_{12}(y, Z_0), S_{21} = f_{21}(y, Z_0)$ and $S_{22} = f_{22}(y, Z_0)$ such that

$$b_1 = S_{11}a_1 + S_{12}a_2 \tag{1.27}$$

$$b_2 = S_{21}a_1 + S_{22}a_2 \tag{1.28}$$

where S_{ij} s are known as scattering parameters or simply S-parameters of the two port network and they are only uniquely defined if the system impedance level is fixed.

The S-parameters are seen to represent reflection or transmission coefficients from above set of equations they are defined as follows

$$S_{11} = \left.\frac{b_1}{a_1}\right| a_2 = 0$$

input reflection coefficient with output properly terminated.

$$S_{12} = \left.\frac{b_1}{a_2}\right| a_1 = 0$$

reverse transmission coefficient with input properly terminated.

$$S_{21} = \left.\frac{b_2}{a_1}\right| a_2 = 0$$

forward transmission coefficient with output properly terminated.

$$S_{22} = \left.\frac{b_2}{a_2}\right| a_1 = 0$$

output reflection coefficient with input properly terminated.

The advantage of using S-parameters is clear from their definitions. They are measured using a matched termination(i,e making $a_1 = 0$ or $a_2 = 0$).Using matched resistive terminations to measure the S-parameters of a transistor has a advantage that the transistor does not oscillate.

1.5 Signal flow graphs

A signal-flow graph is pictorial representation of a system normally described by a set of simultaneous equations. In microwave circuit analysis, circuits are

described in terms of travelling 'power' waves, 'a' s and 'b' s, related each other by S-parameters in the form of linear simultaneous equations. Hence, the signal flow graph technique can be adopted to represent linear microwave circuits pictorially via S-parameters, and furthermore, it can also be used to simplify circuit for analysis.

Certain rules are to be followed in constructing a signal flow graph:

1. Each variable has to be designated as node.
2. The S-parameters and reflection coefficients are represented by branches.
3. Branches enter into the dependent variable nodes and emanate from the independent variable nodes.The independent variable nodes are the incident waves and the reflected waves are dependent variable nodes.

To determine the ratio or transfer function T of dependent to independent variable ,we apply Mason's rule, namely[1]

$$T = \frac{b}{a} = \sum_{k=1}^{n} \frac{M_k \Delta_k}{\Delta}$$

where M_k =product of branches encountered in the 'k'th forward path between nodes 'a' and 'b'.

n =total number of forward paths between 'b' and 'a'.

b =output node variable.

a =input node variable.

$$\Delta = 1 - \sum_m P_{m1} + \sum_m P_{m2} - \sum_m P_{m3} + \dots$$

P_{mr} =product of branches of the m th possible combination of 'r' nontouching loops

$(1 \leq r \leq n)$

Δ_k =the Δ for that part of signal flow graph that is nontouching with k th forward path.

There are some applications are given below

Calculation of input reflection coefficient Γ_{IN}, when a load is connected to the output of a two port network.The signal flow graph is shown in figure 1.2 the input reflection coefficient is defined as

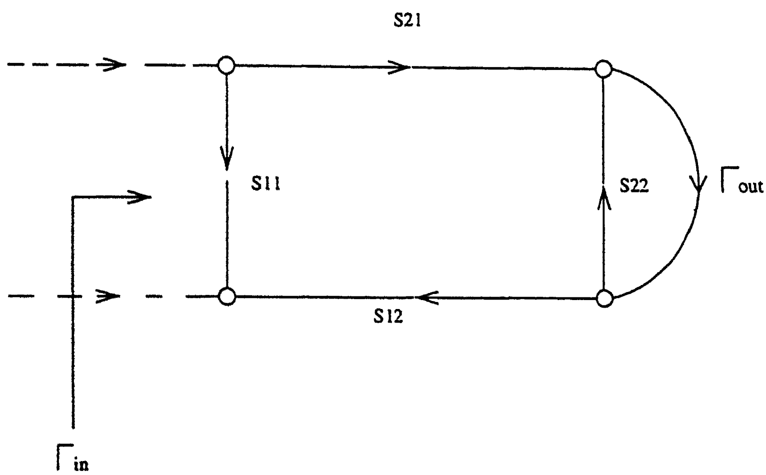

Figure 1.2: Signal flow graph for input reflection coefficient Γ_{IN}

$$\Gamma_{IN} = \frac{b_1}{a_1}$$

by using Mason's rule

$$\begin{aligned}\Gamma_{IN} &= \frac{S_{11}(1 - S_{22}\Gamma_L) + S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \\ &= S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \qquad (1.29)\end{aligned}$$

Calculation of output reflection coefficient Γ_{out}, the signal flow graph is shown in figure 1.3.

$$\Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \qquad (1.30)$$

Similarly signal flow graphs are used in calculation of power gain. There are three different power gains defined as

Transducer power gain[G_T]= $\frac{P_L}{P_{AVS}}$

Operating power gain[G_P]= $\frac{P_L}{P_{IN}}$

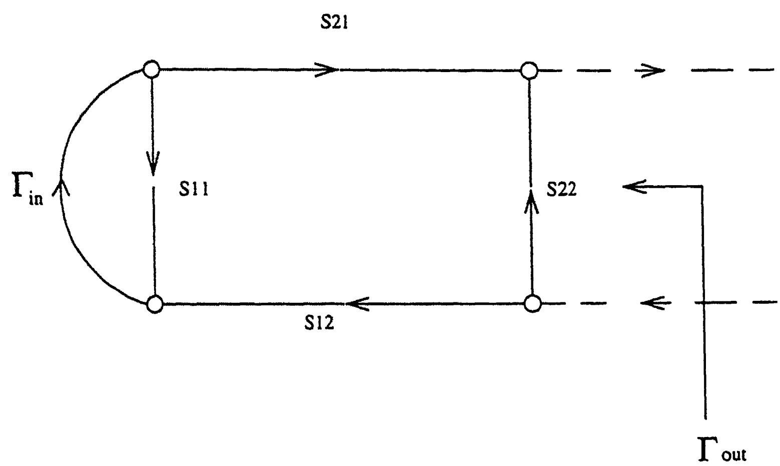

Figure 1.3: Signal flow graph for input reflection coefficient Γ_{IN}

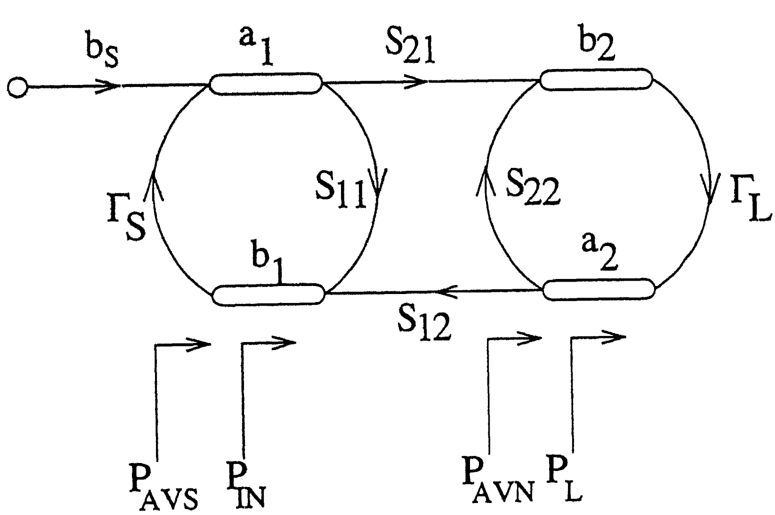

Figure 1.4: Signal flow graph of a microwave amplifier

Available power gain$[G_A]= \frac{P_{AVN}}{P_{AVS}}$

where P_{AVS} is the power available from the source $= P_{IN}$when $\Gamma_{IN} = \Gamma_S^*$

P_{AVN} is the power available from the network $= P_L$when $\Gamma_L = \Gamma_{OUT}^*$

P_{IN} is the power input to the network $= |a_1|^2 - |b_1|^2$

P_L is the power delivered to the load $= |b_2|^2 - |a_2|^2 = |b_2|^2 \left(1 - |\Gamma_L|^2\right)$

we can write the power gain equations in the form

$$G_T = \frac{1 - |\Gamma_S|^2}{|1 - \Gamma_{IN}\Gamma_S|^2} |S_{21}|^2 \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} \tag{1.29}$$

$$G_T = \frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2} |S_{21}|^2 \frac{1 - |\Gamma_L|^2}{|1 - \Gamma_{OUT}\Gamma_L|^2} \tag{1.30}$$

$$G_P = \frac{1}{1 - |\Gamma_{IN}|^2} |S_{21}|^2 \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} \tag{1.31}$$

$$G_A = \frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2} |S_{21}|^2 \frac{1}{1 - |\Gamma_{OUT}|^2} \tag{1.32}$$

These equations will be used in subsequent chapters.

1.6 Organisation of thesis

The thesis has been devided into five chapters.Chapter-1 is introduction which deals with basic concepts in this area and procedures used in other chapters of this thesis. Chapter-2 deals with the explanation of different design considerations like gain, stability, noise figure in the design of microwave amplifier. In chapter-3, design and development procedure has been discussed. In chapter 4 testing, alignment, measurement and results of various parameters are discussed. In the last chapter conclusions and future developments have been discussed.

Chapter 2

Design considerations

The most important design considerations in a narrowband microwave transistor amplifier are stability, power gain, noise and dc requirements.

2.1 Stability considerations

The stability of an amplifier, or its inhibition to oscillate, is the prime consideration in design of amplifier and can be determined from the S-parameters, the matching networks and the terminations. In a two port network, oscillations are possible when either the input or the output port presents a negative resistance. This occurs when $|\Gamma_{IN}| > 1$ or $|\Gamma_{OUT}| > 1$.

The conditions for unconditional stability at a given frequency are

$$|\Gamma_S| < 1 ,$$

$$|\Gamma_L| < 1$$

$$|\Gamma_{IN}| = \left| S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \right| < 1 \tag{2.1}$$

$$|\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \right| < 1 \tag{2.2}$$

where all coefficients are normallised to the same characteristic impedance Z_0.

The graphical analysis is especially useful in the analysis of potentially unstable transistors. When the two port network is potentially unstable, there may be values of Γ_S and Γ_L (i,e source and load reflection coefficients) for

which the real parts of Z_{IN} and Z_{OUT} are positive. These values of Γ_S and Γ_L (i,e regions in the smith chart) can be determined using the following graphical procedure.

First, the regions where values of Γ_S and Γ_L produce $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$ are determined, respectively. Setting the magnitude of expressions 2.1 and 2.2 equal to 1 and solving for the values of Γ_S and Γ_L shows that the solutions for Γ_S and Γ_L lie on the circles (called stability circles) whose equations are given by

$$\left|\Gamma_S - \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2}\right| = \left|\frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2}\right| \tag{2.3}$$

(2.4)

and,

$$\left|\Gamma_L - \frac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2}\right| = \left|\frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2}\right| \tag{2.5}$$

(2.6)

where,

$$\Delta = S_{11}S_{22} - S_{12}S_{21} \tag{2.7}$$

The radii and centers of circles where $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$ inthe Γ_L plane and Γ_S plane, respectively are given by

Γ_L values for $|\Gamma_{IN} = 1|$ (output stability circle):

$$r_L = \left|\frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2}\right| \text{(.....radius)} \tag{2.8}$$

$$C_L = \frac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2} \text{(.....center)} \tag{2.9}$$

(2.10)

Γ_S values for $|\Gamma_{OUT}| = 1$ (Input stability circle):

$$r_S = \left|\frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2}\right| \text{(.....radius)} \tag{2.11}$$

$$C_S = \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} \text{(.....center)} \tag{2.12}$$

With the S-parameters of a two port device at one frequency, the centers and radii can be calculated, plotted on a smith chart, and the set of values of Γ_L and Γ_S that produce $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$ can be easily observed. Figure 2.1 illustrates the graphical construction of the stability circles where $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$. On one side of stability circle boundary, in the Γ_L plane, we will have $|\Gamma_{IN}| < 1$ and on the other side $|\Gamma_{IN}| > 1$. Similarly , in the Γ_S plane, on one side of stability circle boundary we will have $|\Gamma_{OUT}| < 1$ and on the other side $|\Gamma_{OUT}| > 1$.

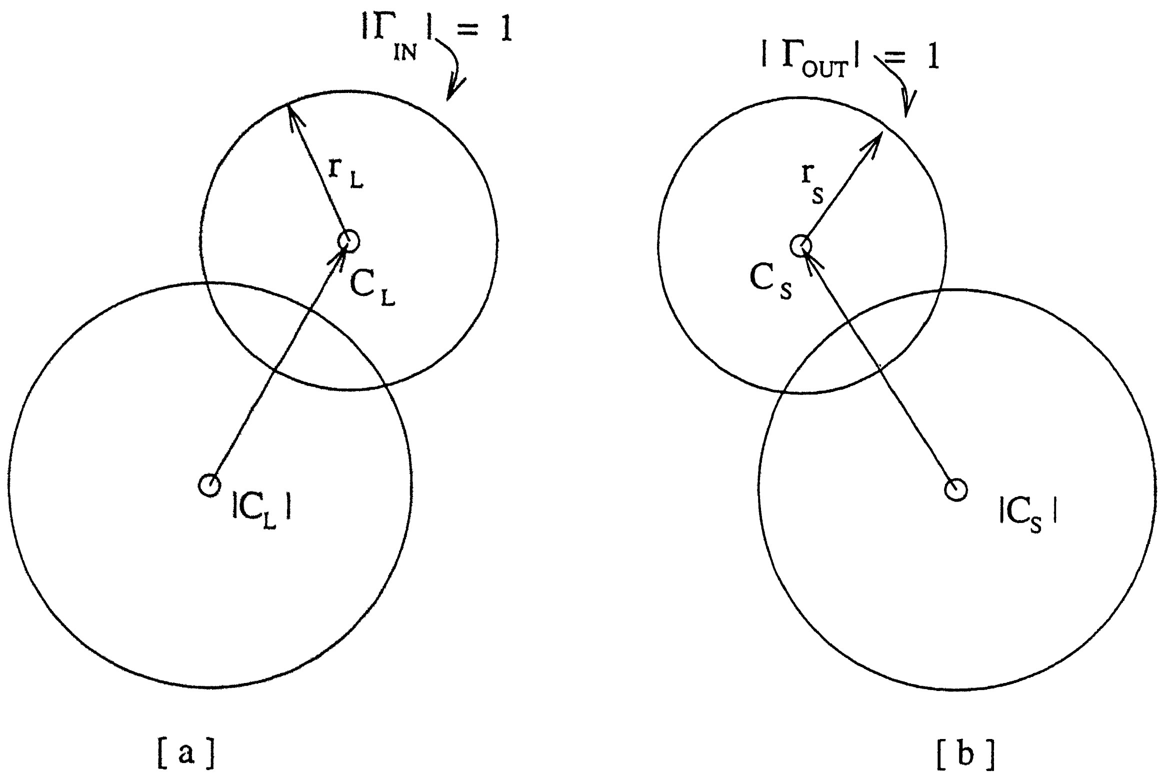

Figure 2.1: Stability circles construction in the smith chart:[a] Γ_L plane; [b] Γ_S plane.

Next, we need to determine which area in the smith chart represents the stable region. In other words, the the region where values of Γ_L produce $|\Gamma_{IN}| < 1$, and where Γ_S produce $|\Gamma_{OUT}| < 1$. We observe that if $Z_L = Z_0$, then $\Gamma_L = 0$ and $|\Gamma_{IN}| = |S_{11}|$. If magnitude of S_{11} is less than 1, then $|\Gamma_{IN}| < 1$ when $\Gamma_L = 0$. That is center of the smith chart in Figure 2.1 represents a stable operating point. On the otherhand, if $|S_{11}| > 1$ when $Z_L = Z_0$, then $|\Gamma_{IN}| > 1$ when $\Gamma_L = 0$ and the center of the smith chart represents an unstable

operating point. Figure 2.2 illustrates the two cases discussed. The shaded area represents the values of Γ_L that produce a stable operation. Similarly Figure 2.3 illustrates stable and unstable regions for Γ_S.

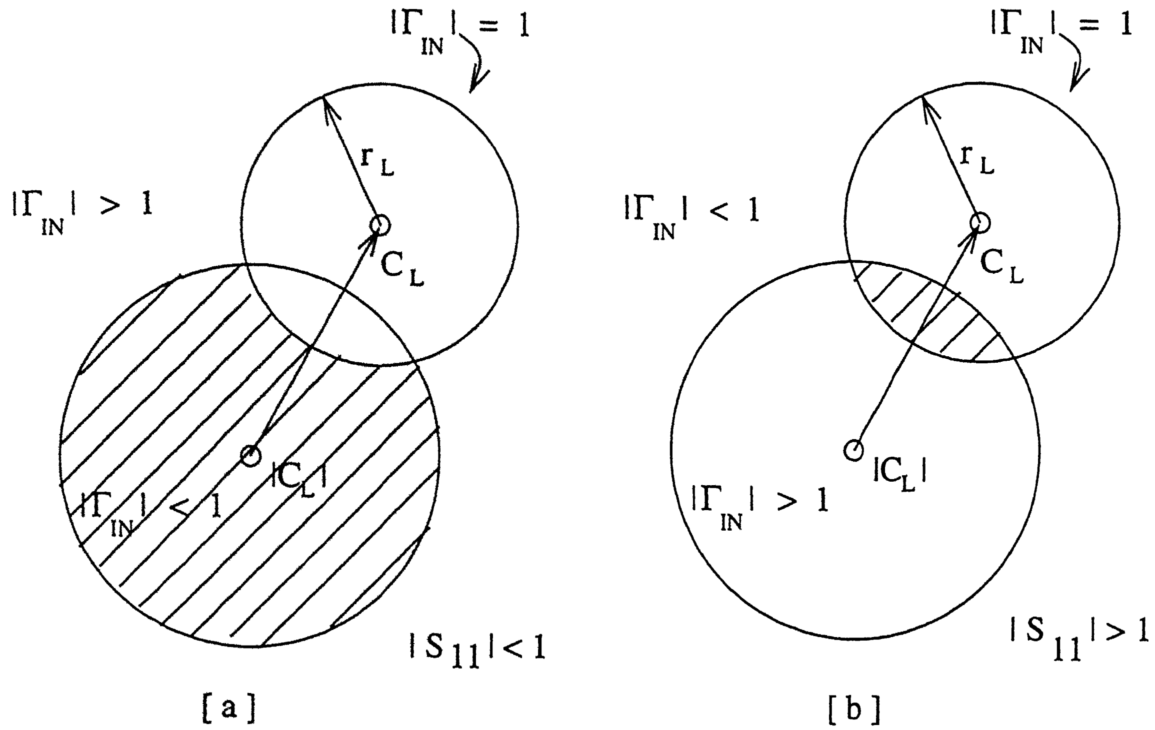

Figure 2.2: Smith chart illustrating stable and unstable regions in the Γ_L plane.

The necessary and sufficient conditions for a two port network to be unconditionally stable are given by

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2\,|S_{12}S_{21}|} > 1 \tag{2.13}$$

$$\text{and } |\Delta| < 1 \tag{2.14}$$

where

$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}|$$

In conclusion, when $|S_{11}| < 1$ and $|S_{22}| < 1$, the two port network is unconditionally stable if $K > 1$ and $|\Delta| < 1$.

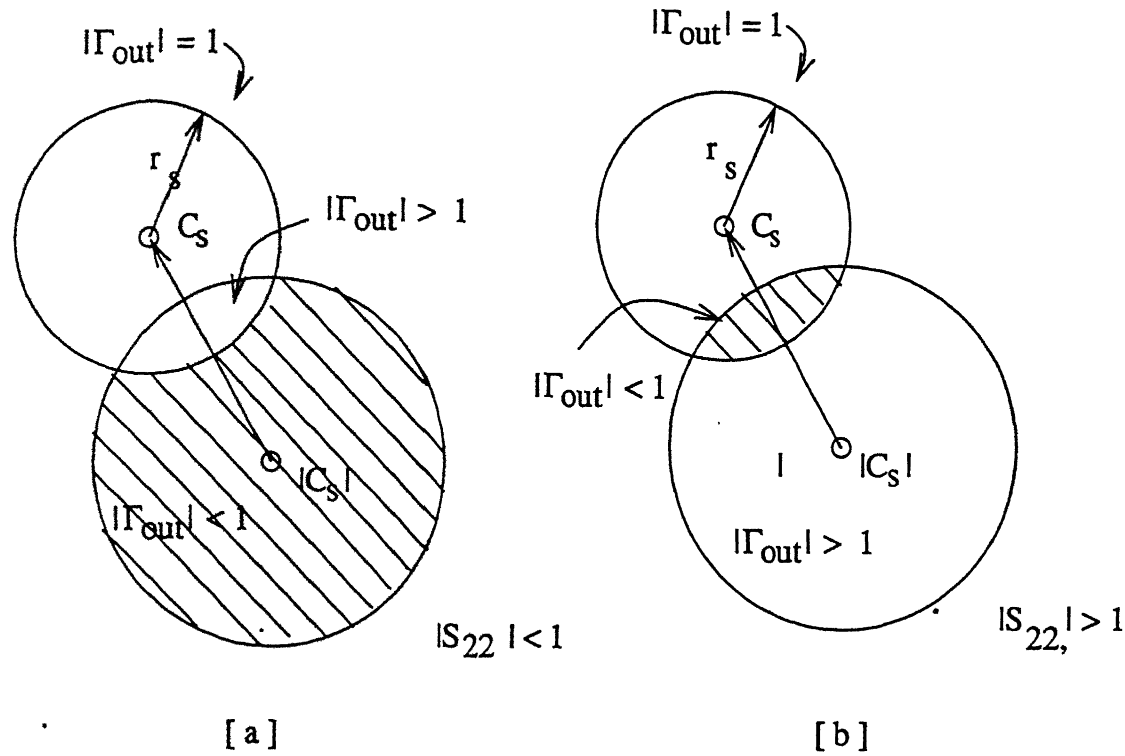

Figure 2.3: Smith chart illustrating stable and unstable regions in the Γ_S plane.

2.2 Available power gain circles

We can use either operating gain circles or available gain circles for design purpose. The constant available gain circles and constant noise figure circles are functions of Γ_S and they can be plotted together on the smith chart (i,e Γ_S plane) and variation of gain and noise figure parameters at various points in the Γ_S plane can be analysed . So, Design procedure based on available gain, G_A, is commonly used. The available power gain is independent of the load impedance, therefore, an available power gain circle procedure for both unconditionally stable and potentially unstable transistors is simple and recommonded for practical designs.

To develop design procedure with G_A, we write the expression 1.32 in the form of

$$G_A = \frac{|S_{21}|^2\left(1-|\Gamma_S|^2\right)}{\left(1-\left|\frac{S_{22}-\Delta\Gamma_S}{1-S_{11}\Gamma_S}\right|^2\right)|1-S_{11}\Gamma_S|^2}$$
$$= |S_{21}|^2 g_a \tag{2.15}$$

where

$$g_a = \frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2 - |S_{22}-\Delta\Gamma_S|^2}$$
$$= \frac{1-|\Gamma_S|^2}{1-|S_{22}|^2+|\Gamma_S|^2\left(|S_{22}|^2-|\Delta|^2\right)-2Re\left(\Gamma_S C_1\right)} \tag{2.16}$$

where

$$C_1 = S_{11} - \Delta S_{22}^*$$

Here G_A and g_a are the functions of S-parameters and Γ_S

The constant available power gain circles are obtained by letting $\Gamma_S = U_S + jV_S$ and substituting into expression 2.16. we can write as

$$\left[U_S - \frac{g_a Re\left[C_1^*\right]}{1+g_a\left(|S_{11}|^2-|\Delta|^2\right)}\right]^2 + \left[V_S - \frac{g_a Im\left[C_1^*\right]}{1+g_a\left(|S_{11}|^2-|\Delta|^2\right)}\right]^2$$
$$= \left[\frac{\left[1-2K|S_{12}S_{21}|g_a+|S_{12}S_{21}|^2 g_a^2\right]^{1/2}}{1+g_a\left(|S_{11}|^2-|\Delta|^2\right)}\right]^2 \tag{2.17}$$

Expression 2.17 is recognised as a family of circles in the U_S, V_S plane (i,e the smith chart) with g_a as the parameter. The center of the circle is located at

$$U_a = \frac{g_a Re\left[C_1^*\right]}{1+g_a\left(|S_{11}|^2-|\Delta|^2\right)} \tag{2.18}$$
$$V_a = \frac{g_a Im\left[C_1^*\right]}{1+g_a\left(|S_{11}|^2-|\Delta|^2\right)} \tag{2.19}$$

The radius of the circle is given by

$$R_a = \frac{\left[1-2K|S_{12}S_{21}|g_a+|S_{12}S_{21}|^2 g_a^2\right]^{1/2}}{1+g_a\left(|S_{11}|^2-|\Delta|\right)} \tag{2.20}$$

The distance from the origin of the smith chart to the centers of the circles is given by

$$d_a = \sqrt{U_a^2 + V_a^2} = \frac{g_a C_1^*}{\left|1 + g_a \left(|S_{11}|^2 - |\Delta|^2\right)\right|} \tag{2.21}$$

Therefore centers of circles can be written as

$$C_a = d_a$$

The maximum available power gain occurs when $R_a = 0$. Therefore, from expression 2.20 we can write

$$1 - 2K\,|S_{12}S_{21}|\, g_{amax} + |S_{12}S_{21}|^2\, g_{amax}^2 = 0 \tag{2.22}$$

where g_{amax} is the maximum value of g_a. The solution for the above equation for unconditional stability is

$$g_{amax} = \frac{1}{|S_{12}S_{21}|}\left(K - \sqrt{K^2 - 1}\right) \tag{2.23}$$

therefore

$$G_{Amax} = \frac{|S_{21}|}{|S_{12}|}\left(K - \sqrt{K^2 - 1}\right) \tag{2.24}$$

The procedure for drawing a constant available gain circles in the Z-Smith chart is as follows.

1. For a given G_A, the radius and center of constant available gain circles are to be calculated.

2. Select the desired Γ_S.

3. For the given Γ_S, maximum output power is obtained with conjugate match at the output namely with $\Gamma_L = \Gamma_{OUT}^*$, where Γ_{OUT} can be calculated from Γ_S. This values of Γ_S produces the transducer power gain G_T.

2.3 Constant noise figure circles

The noise figure of two port amplifier is given by

$$F = F_{min} + \frac{4r_n \left|\Gamma_S - \Gamma_{opt}\right|^2}{\left(1 - \left|\Gamma_S\right|^2\right) \left|1 + \Gamma_{opt}\right|} \tag{2.25}$$

where r_n is the equivalent normalized noise resistance, Γ_{opt} is the source reflection coefficient, which results in minimum noise figure F_{min}.

The equation 2.25 depends on F_{min}, r_n and Γ_{opt}. These quantities are known as the noise parameters and are given by the manufacturer of the transistor or can be determined experimentally. The source reflection coefficient can be varied until a minimum noise figure is read in the noise figure meter. The source reflection coefficient that produces F_{min} can be determined accurately using a network analyser. The noise resistance can be measured by reading noise figure when $\Gamma_S = 0$, called $F_{\Gamma_S=0}$. Then from equation 2.25 we obtain

$$r_n = \left(F_{\Gamma S=0} - F_{min}\right) \frac{\left|1 + \Gamma_{opt}\right|^2}{4\left|\Gamma_{opt}\right|^2} \tag{2.26}$$

Equation 2.25 can be used to design Γ_S for a given noise figure. For a given noise figure F_i, we define a noise figure parameter, called N_i, as

$$N_i = \frac{\left|\Gamma_S - \Gamma_{opt}\right|^2}{1 - \left|\Gamma_S\right|^2} = \frac{F_i - F_{min}}{4r_n} \left|1 + \Gamma_{opt}\right|^2 \tag{2.27}$$

the equation 2.27 can be written as

$$\left(\Gamma_S - \Gamma_{opt}\right)\left(\Gamma_S^* - \Gamma_{opt}^*\right) = N_i - N_i \left|\Gamma_S\right|^2 \tag{2.28}$$

or

$$\left|\Gamma_S\right|^2 (1 + N_i) + \left|\Gamma_{opt}\right|^2 - 2Re\left(\Gamma_S \Gamma_{opt}^*\right) = N_i \tag{2.29}$$

or

$$\left|\Gamma_S - \frac{\Gamma_{opt}}{1 + N_i}\right|^2 = \frac{N_i^2 + N_i\left(1 - \left|\Gamma_{opt}\right|^2\right)}{(1 + N_i)^2} \tag{2.30}$$

The equation 2.30 is of the form of circles with N_i as parameter. The circles are centered at

$$C_{F_i} = \frac{\Gamma_{opt}}{1 + N_i} \tag{2.31}$$

with radii,

$$R_{F_i} = \frac{1}{1+N_i}\sqrt{N_i^2 + N_i\left(1 - |\Gamma_{opt}|^2\right)} \tag{2.32}$$

The center of the F_{min} circle is located at Γ_{opt} with zero radius, the centers of the other noise figure circles are located along the Γ_S vector.

In a design there is always a difference between the designed noise figure and measured noise figure of the final amplifier. This occurs because of the loss associated with the matching elements and the transistor noise figure variations from unit to unit.

2.4 DC bias considerations

There are different bias circuit configurations depending on the application like high power, or low power, or low noise figure. Generally, in GaAsFET for low power applications source resistor bias arrangement is prefered. In this type, source resistor degrades the noise figure performance and the source by pass capacitor can cause low frequency oscillations.

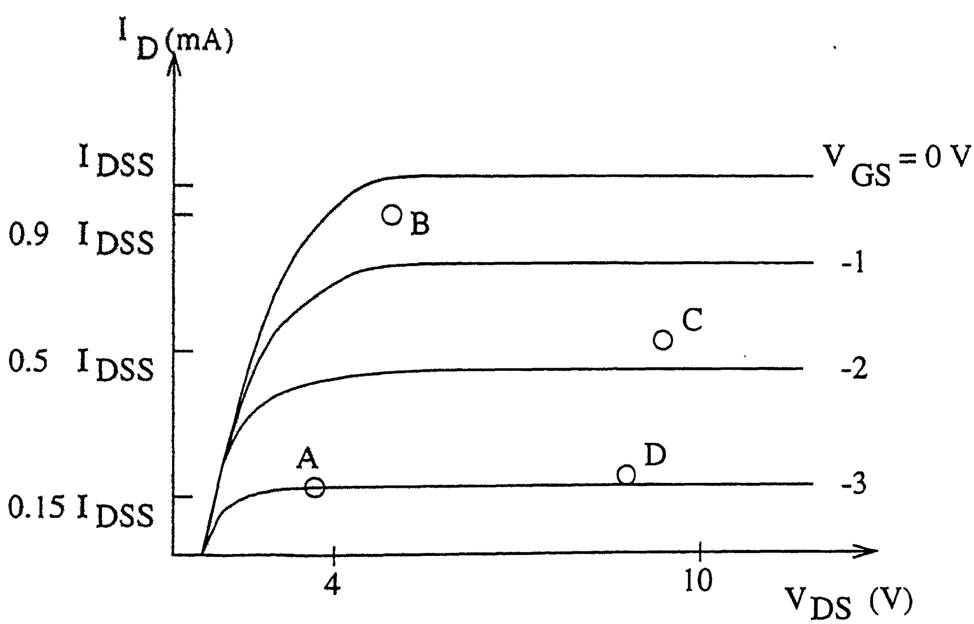

Figure 2.4: Typical GaAs FET characteristics and recommonded quiescent points.

The dc bias network of GaAs FET must provide a stable quiescent point. The selection of quiescent point in GaAs FET depends on the particular appli-

cation. Figure 2.4 shows typical GaAs FET characteristics with four quiescent points located at A, B, C and D [1].

For low noise, low power application,the quiescent point A is recommonded. At A, the FET operates at a low value of current(i,e $I_{DS} \approx 0.15 I_{DSS}$).

For low noise and slightly higher power gain, the recommonded point is at B. The bias voltage remains the same as point A, but the drain current is increased to $I_{DS} \approx 0.9 I_{DSS}$.

The GaAs FET output power level can be increased by selecting the quiescent point at C with $I_{DS} \approx 0.5 I_{DSS}$. The quiescent point maintains class A operation. For higher efficiency, or to operate the GaAs FET in class AB or B, the drain-to-source current must be decreased and the quiescent point D is recommonded.

Chapter 3

Design and development of microwave transistor amplifier

In previous chapters we have discussed various basic considerations required for the design of a microwave transistor amplifier. Based on these theory, we can conclude that there are three major steps are involved in designing an amplifier. We can devide into three separate sections in designing.

1. Design of biasing network.

2. Selection of source reflection coefficient.

3. Design of input and output matching networks.

3.1 Choice of transistor and its biasing

In the design of any active circuit, choice of transistor (or any other active element) is very critical. Because it decides the maximum available gain and noise figure of the circuit.

So we should select a transistor of desired capability and frequency range. A FET NE.70083[NEC] is selected for this amplifier design with the following specification at 8 GHz.

Optimum noise figure $(NF_{opt}) = 1.2dB$

Maximum available gain $(G) = 16dB$

Associated Gain at $F_{opt}\,(G_A) = 11dB$

and S-parameters at $V_{ds} = 3volts$ and $I_{ds} = 10mA$ are

$$S_{11} = 0.77\angle 124$$

$$S_{12} = 0.07\angle 6$$

$$S_{21} = 1.83\angle 54$$

$$S_{22} = 0.66\angle -33$$

Biasing of the transistor Design of biasing circuit depends on choice of suitable quiescent point on device characteristics as discussed in chapter 2. characteristic depending on the application as discussed in chapter 2. Figure 3.1 shows typical circuit diagram of an amplifier. The values of R, L_1, L_2, C_c and C_b are calculated as follows.

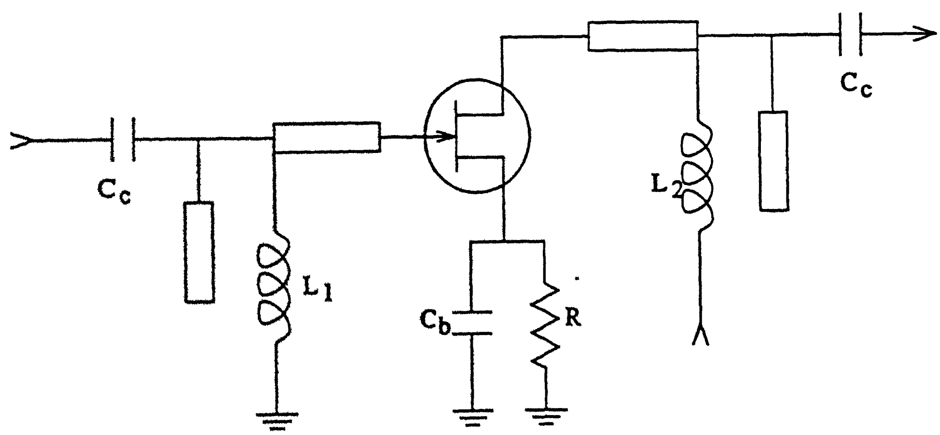

Figure 3.1: Circuit diagram.

C_c and C_b must act as short circuit at 8.0 GHz and as open circuit at DC. Therefore,

$\frac{1}{wc} \approx 0.2$ (say) at 8.0 GHz

$\frac{1}{wc} \approx \infty$ at DC.

$C_c and C_b \approx 0.1nF = 100pF$

Inductors (L_1 and L_2) should be chosen in a manner that it act as short circuit at DC and open circuit at 8.0 GHz. Also no output signal should leak to the ground.

R can be calculated as follows

$i = 10mA$

$V_{ds} = 3volts$

$V_{dd} = 3.6volts$

$V_{dd} = V_{ds} + V_s$

$R = \frac{V_s}{i} = 60\Omega$

3.2 Selection of source reflection coefficient

Selection of source reflection coefficient places a vital role in design of a microwave amplifier. One can choose either Γ_S or Γ_L. Generally Γ_S is chosen One can choose either Γ_S or Γ_L. But we are choosing Γ_S as gain and noise figure circles are being plotted on Γ_S plane. Then Γ_L can be calculated from Γ_S. Figure 3.2 shows the block diagram of a microwave amplifier.

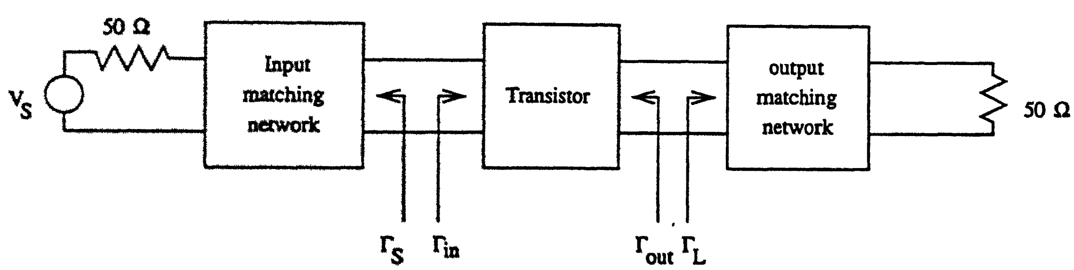

Figure 3.2: Block diagram of microwave amplifier.

There are four steps involved in the selection of Γ_S in Γ_S- plane.

1. The point Γ_S should be in the stable region, and also Γ_L calculated from Γ_S should also be in the stable region.

2. The point Γ_S should lie on the specified gain circle.

3. The point Γ_S should lie on the specified noise figure circle.

4. The point Γ_S should result in a compromise value of VSWR.

For a given S-parameters, gain, noise figure, optimum noise figure, and noise resistance, we need to calculate the values of K, Δ , centers and radii of source stability-, load stability-, gain and noise figure circles. From the values of K

and Δ, the stability of the two port network can be decided. If the two port network is unstable, then there are two possible ways to design. First one is to make the network unconditionally stable by employing feed back network and second one is not to select Γ_S point in the unstable region in the Γ_S-plane.

In order to satisfy the specified gain and noise figure requirements Γ_S point should be on the gain and the noise figure circles which in terms will give one of the intersecting points of the two circles. That intersecting points should be far from unstable region and should give a reasonable VSWR.

By using so chosen Γ_S, Γ_{OUT} and from which Γ_L can be calculated. But one should make sure that Γ_L is also in the output stable region. Those values of Γ_S and Γ_L can be used in design of input and output matching networks.

3.3 Design of input and output matching networks

Matching networks play crucial role on the frequency response of a microwave amplifier. For narrowband applications single stub matching is prefered. Manually, design can be done using smith chart or using formulas. Single stub matching using smith chart method is programmed and calculated results through program are given in next section.

3.4 Software description

Software has been developed for the design of narrowband microwave amplifier. Figure 3.3 shows the basic flow chart of software. The flow diagram explains how the program makes design of narrowband microwave amplifier. First program reads the data file, which consists all specifications and parameters of transistor. It makes calculation of necessary parameters for selection of Γ_S point. It gives option for selection of Γ_S point, which is explained in features. With this Γ_S and corresponding Γ_L it designs a single stub matching networks for source and load sides. After that it calculates width, wavelength and other parameters of microstrip by using synthesis formulae given in chapter 1. Fi-

nally it makes a data file, which creates PCB layout.

Some important features in software are given below.

1. By giving the S-parameters, F_{min}, Γ_{opt}, R_n, required gain and noise figure in dBs, ϵ_r, thickness of dielectric material, characteristic impedance of stub and series transmission lines, frequency of operation and transistor dimensions to the input data file, this software gives a data file, which is data for PCB layout, and creates an output file, which consists of the values of different parameters like K, Δ, centers of various circles, Γ_S, Γ_L, Y_S, Y_L, width of microstrip, λ_g, VSWR, Transducer gain etc.

2. If the given gain and noise figure requirement values do not give the solution for Γ_S point, this software makes design for nearest possible gain or noise figure by keeping either gain or noise figure fixed as per our requirement. Thus, trial and error feeding of gain and noise figure input values, until those circle crossover, is avoided.

3. This software selects a better Γ_S point by keeping good compromise between noise figure, gain and VSWR.

4. Normally, in single stub matching network design using smith chart, there will be two points (i.e., intersection of reflection coefficient circle and constant conductance circle). Some times Γ_S point may fall very nearer to constant conductance circle, which results in very low series length of single stub matching network(i.e., which can not be physically realised at lower wavelengths). This software calculates series line and shunt stub lengths of single stub matching network for both the points and gives option for selection of one point, in which strip lengths can be physically realised.

5. Finally, after design of matching network, this software gives an output file consisting of data, which creates PCB layout.

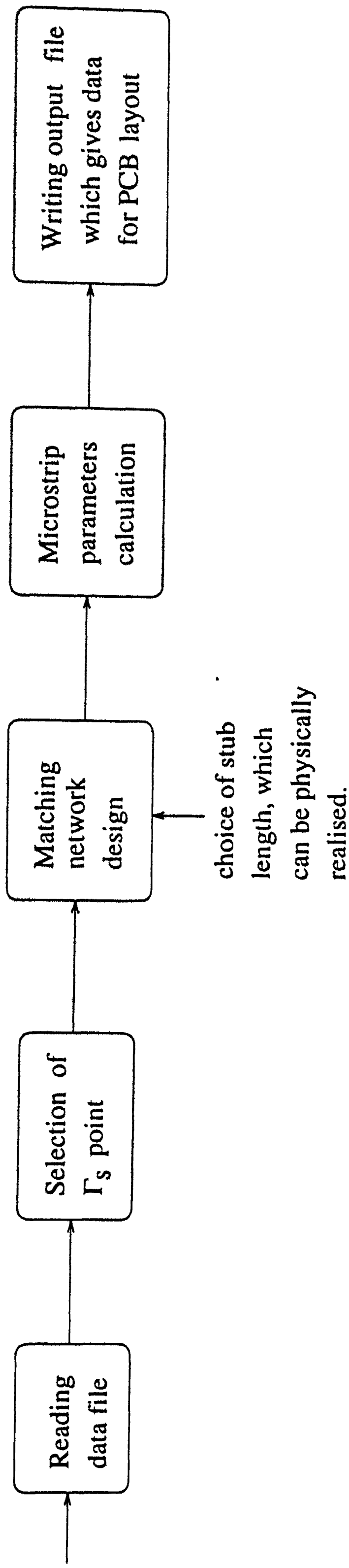

Figure 3.3: Flow diagram of software

Input data

S-parameters	: S_{11}	**0.77∠-124**
	S_{12}	**0.07∠6**
	S_{21}	**1.93∠54**
	S_{22}	**0.66∠-93**
Noise parameters	F_{min}	1.2dB
	Γ_{opt}	0.62∠118
	R_n	0.0044Ω
Microstrip parameters	: Z_0	50Ω
	: ϵ_r	2.23
	hight	0.7874mm
	Z_{stub}	50Ω
Specification requirements	: gain	13.7dB
	noise figure	1.21dB
	operating frequency	8 GHz
Transistor dimensions	: length	4.5mm
	width	4.0mm
Layout dimensions	: length	120mm
	width	80mm

Output data

Stability measures	: $\|\Delta\|$	0.5097
	K	0.8560
Input stability circle	: center	$1.3634\angle 135.23$
	radius	0.5056
Output stability circle	: center	$1.7047\angle 113.15$
	radius	0.7684
Gain circle	: center	$0.9258\angle 135.23$
	radius	0.1659
Noise figure circle	: center	$0.5479\angle 118$
	radius	0.2771
Input reflection coefficient	: Γ_{IN}	$0.9169\angle 225.24$
Output reflection coefficient	: Γ_{OUT}	$0.8050\angle 252.2$
Source reflection coefficient	: Γ_S	$0.7614\angle 133.72$
Load reflection coefficient	: Γ_L	$0.8050\angle 107.79$
Source admittance	: Y_S	$0.7973\angle$-2.08
Load admittance	: Y_L	$0.3045\angle$-1.32
Input matching network	: stub length	$0.1859\lambda_g$
	stub width	2.4185mm
	series length	$0.1204\lambda_g$
Output matching network	: stub length	$0.1938\lambda_g$
	stub width	2.4185mm
	series length	$0.1508\lambda_g$
Microstrip parameters	: width(w)	2.4185mm
	thickness(h)	0.7874mm

	ϵ_{eff}	1.8686
	$\epsilon_{eff}(f)$	1.8964
	wavelength (λ_g)	27.23mm
Specification parameters	: available gain	13.7dB
	transducer gain	13.84dB
	noise figure	1.21dB
	input VSWR	1.9
	output VSWR	1.0

3.5 Fabrication

After completion of PCB design, microstrip layout is shaped on duroid(ϵ_r = 2.23) sheet with copper coat. FET and other components are soldered properly. Whole layout is fixed in metal enclosure and connectors are soldered. This completes the fabrication of the narrowband microwave amplifier. PCB layout is shown in figure 3.4.

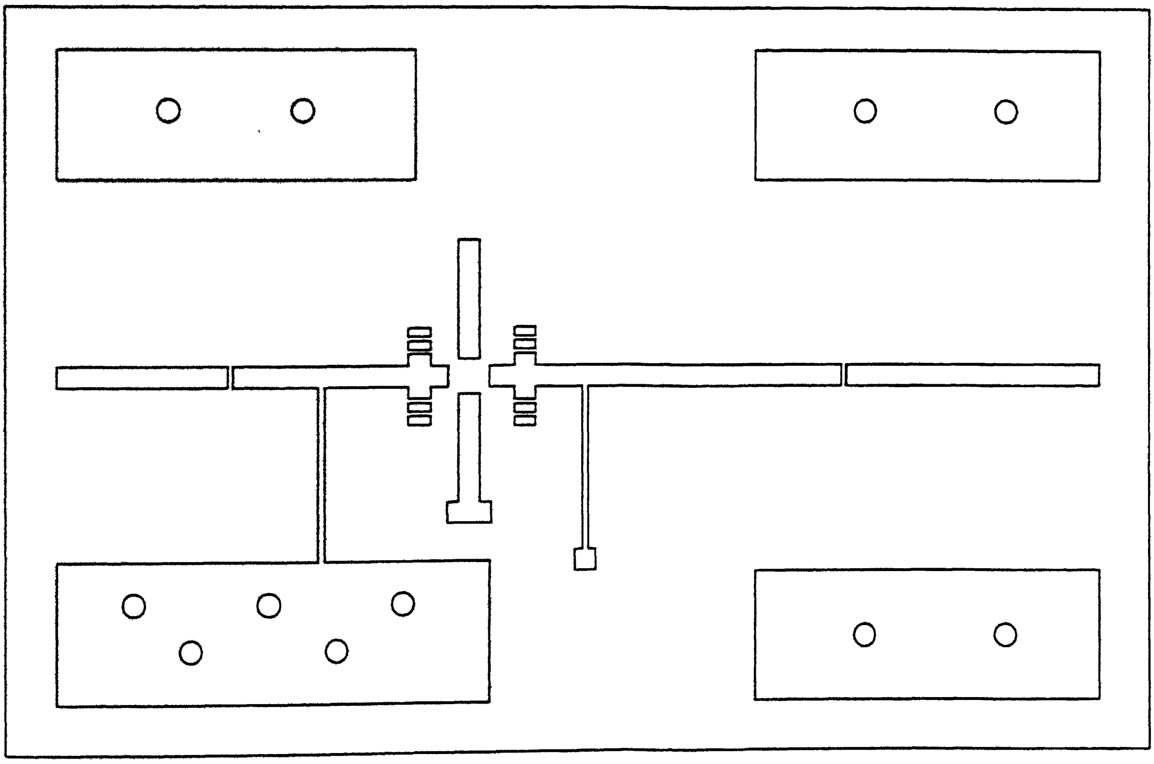

Figure 3.4: PCB layout

Chapter 4

Testing, Alignment and Results

4.1 Testing

Amplifier is designed to operate with 3.6 Volts DC source. Figure shows the test setup for an amplifier circuit. In theoritical design we have chosen Γ_S and Γ_L to fall in the stable region, accordingly calculations of other parameters are done, to make the device as an amplifier. But in practical, there are chances that device may work as an oscillator, even though designed for an amplifier. This is because of change in parameters provided in manual from device to device and accuracy of fabrication etc.. So first we have to test if the device is working as an amplifier or giving any oscillations. Figure 4.1 shows the test setup. Spectrum analyser is set in track mode to view the large band. Microwave source is connected at the input of amplifier, initially microwave source is kept in OFF position. DC supply to the amplifier is switched on and observed for oscillations (if any). It is found that device is oscillating at 6.8 GHz.

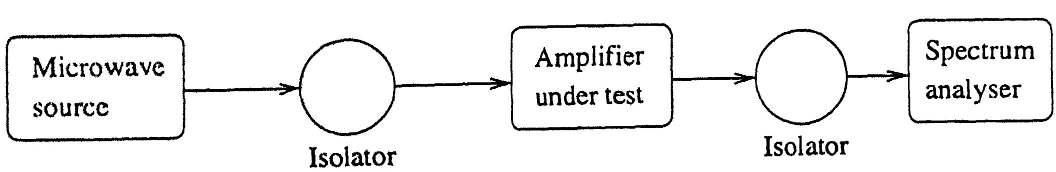

Figure 4.1: Test setup

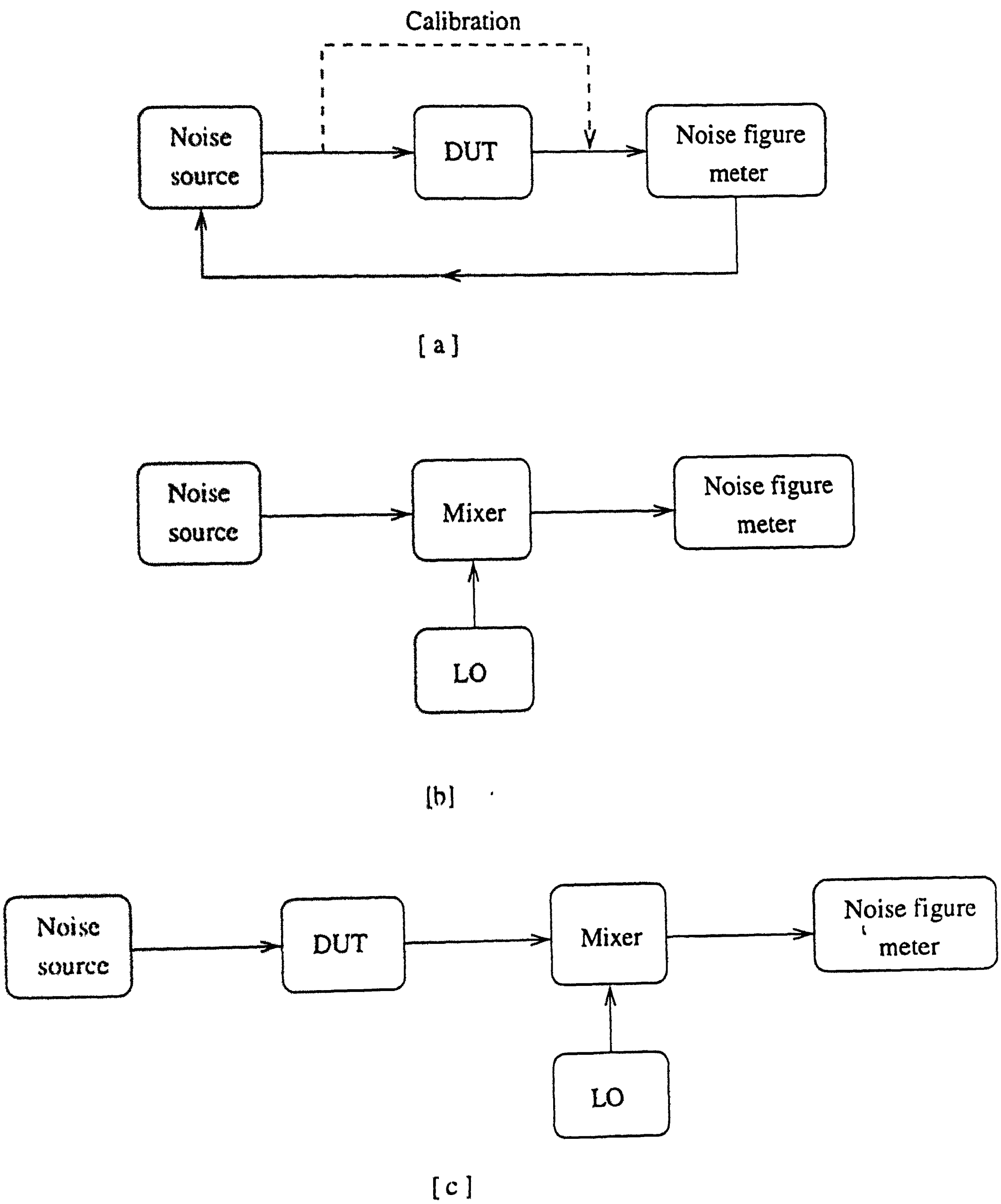

Figure 4.2: Noise figure measurement

4.2 Alignment

Alignment is carried out by taking copper strip, which acts as stub. This stub is moved on the microstrip transmission line at the input side of the device, by observing spectrum analyser, until oscillations dies down. Then the stub is soldered, at which there are no oscillations. Microwave source is switched on and 0 dBm power is fed to input of the amplifier through isolator. It is found that amplifier is working at 6.85 GHz with gain of 14.1 dB.

4.3 Measurements and Results

Figure 4.2 shows the noise figure and gain measurement setup. Normally, noise figure meter operating range is from 10-1600 MHz. Within this range we can measure the parameters, after calibration, directly by feeding the noise source to the input of the DUT and taking output from the DUT and shown in figure 4.2(a). If our operating frequency crosses the above range, we need to use a mixer and LO to bring down the frequency. The complete measurement setup is shown in figure 4.2(b) and (c). The setup is first calibrated without DUT. Then amplifier is connected to it for direct measurement of gain and noise figure. By changing LO frequency, readings of gain and noise figure are taken and are plotted. The gain and the noise figure responses are shown in figure 4.3 and 4.4.

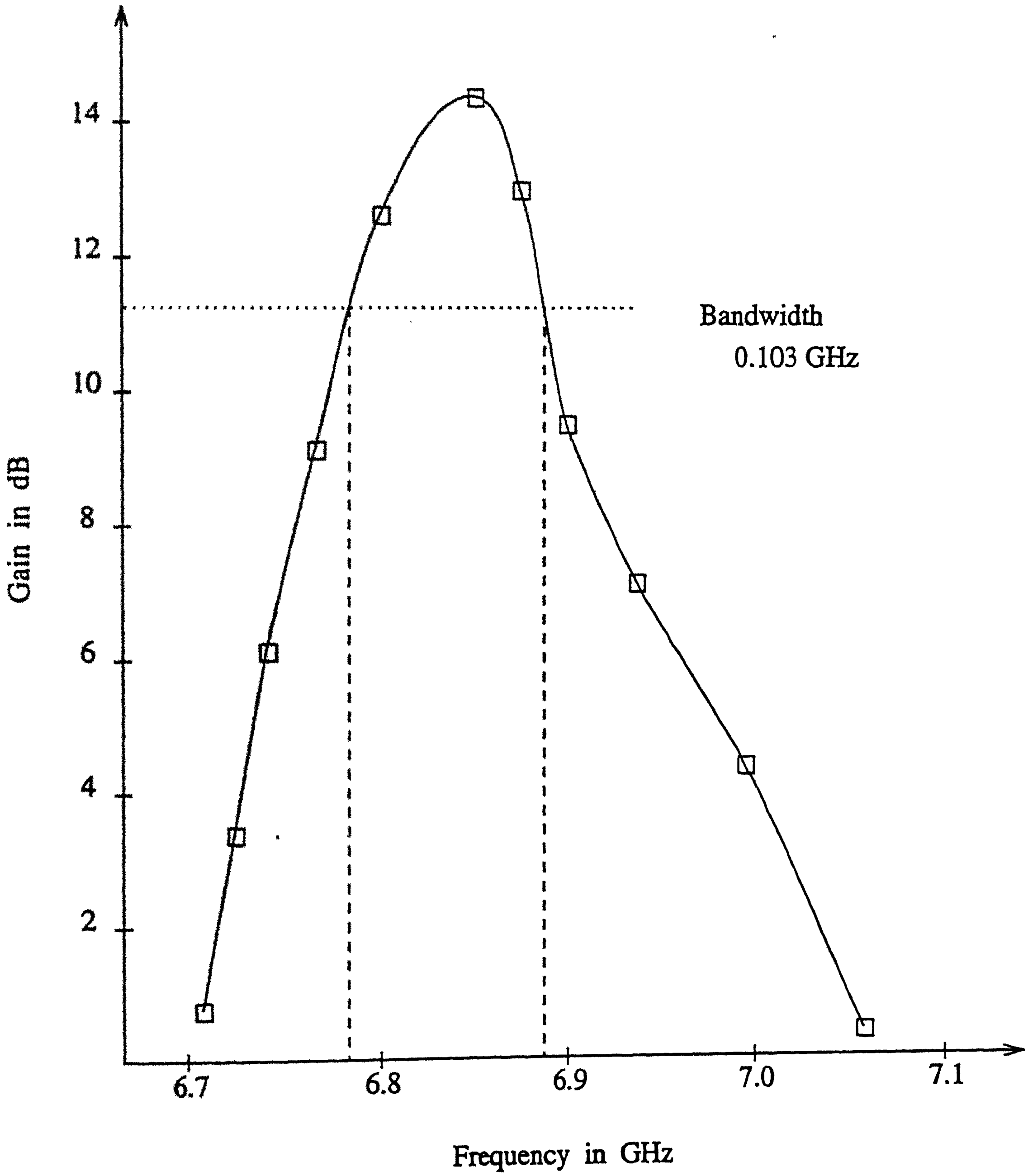

Figure 4.3: Variation of gain with frequency

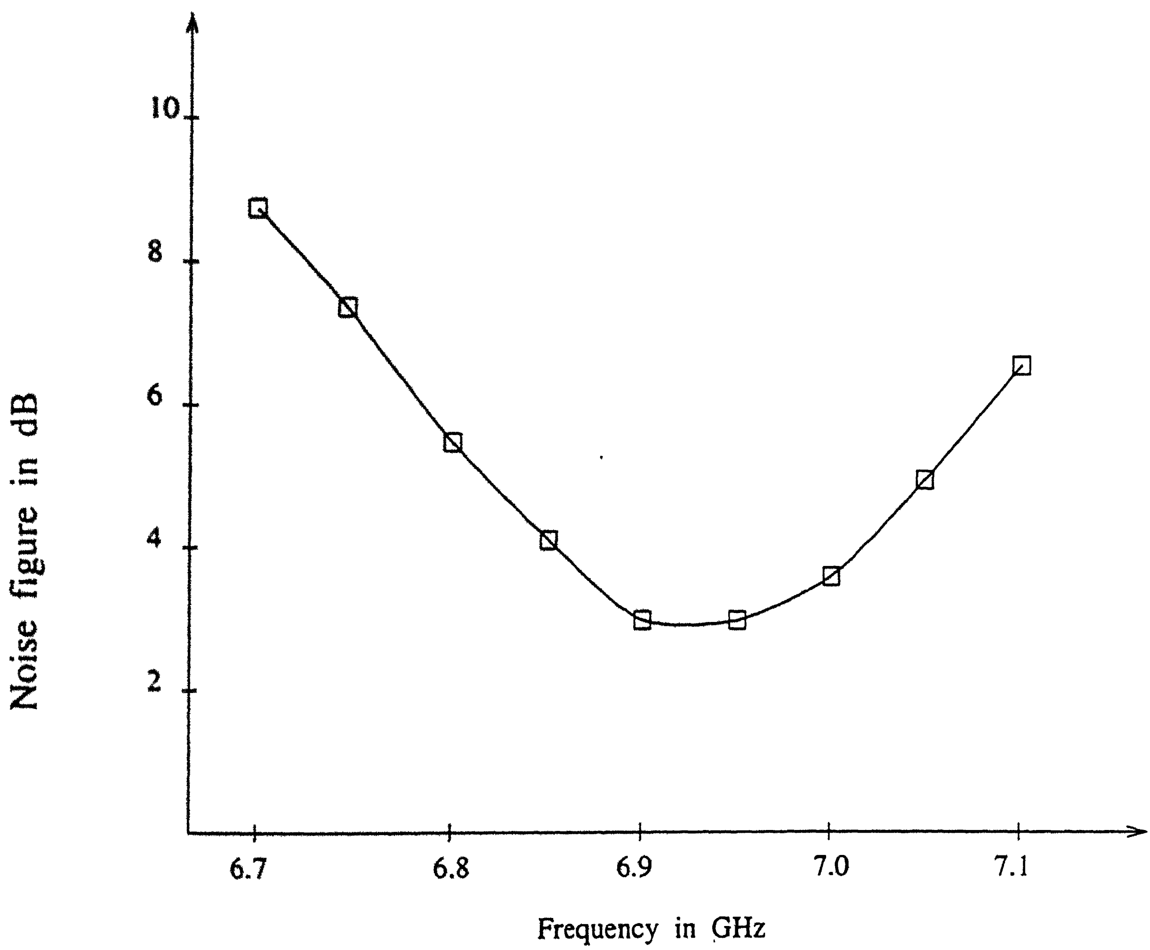

Figure 4.4: Noise response

Chapter 5

Conclusions and scope of future work

5.1 Conclusions

In the design of an amplifier matching networks and selection of Γ_S plays an important role. Various transition effects causes improper matching which causes change of noise figure of the device, as it is dependent on Γ_S. Soldering job has to be carried out very carefully, to avoid transition effects from leads of components to microstripline.

Generally, we are choosing $\Gamma_L = \Gamma^*_{OUT}$ for maximum output power (i,e gain). It gives output VSWR= 1. Sometimes this type of selection may result in high input VSWR. Then by slightly shift in Γ_L from Γ^*_{OUT} gives better input VSWR.

In this thesis work, various design considerations like stability, gain, noise figure, DC bias, matching networks etc. in the design of a microwave transistor amplifier have been studied and practically implemented in laboratory. Various tests have been conducted on the device and found that it is working satisfactory.

Software has been developed for all calculations and for decision making at various steps in design. This software can be used for complete design of narrowband microwave amplifier. This software can also be used for calculations in design of broadband amplifiers and oscillators.

5.2 Scope of Future work

1. Design can be extended to broadband amplifier, by considering Γ_S at different frequencies through out the band and designing broadband matching network.
2. By choosing Γ_S point in unstable region, making $|\Gamma_{IN}|$ much greater than 1 and making load matching network resonates Z_{IN}, oscillator can be designed.
3. Software can be extended for the above designs.

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